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Linearization Trade-Offs in a 5G mmWave Active Phased Array OTA Setup

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ABSTRACT The new generation of 5G mobile communication systems is using millimeter wave (mmWave) active phased arrays (APA) which have up to hundreds of individual analog transmitter and receiver chains and antennas. For these highly integrated systems linearization of each analog path is very challenging. Therefore a single input single output (SISO) system in combination with over the air (OTA) measurement is considered as an efficient approach for linearization. However, the knowledge about the dependency of the total SISO nonlinearity on the contributions from different blocks in the antenna array, as well as the linearization trade-offs is still missing. In this paper, an overview of the possible linearization trade-offs in an OTA setup with a mmWave APA is provided. The linearization technique is applied to a 4×4 active phased array containing up-conversion of a sub 6 GHz LTE10 signal to an RF frequency of 28 GHz. Through measurements, the effects on adjacent channel power ratio (ACPR) and error vector magnitude (EVM) have been investigated for the following scenarios: i. impact from the up-converter, ii. impact of the steering angle due to antenna crosstalk and iii. a linearity comparison between a linearized and a backed-off system.

INDEX TERMS Active phased array (APA), single input single output (SISO), over the air (OTA), power amplifier (PA), millimeter wave (mmWave), digital pre-distortion (DPD).

I. INTRODUCTION

For modern communication systems, high power efficiency is required while maintaining linear operation to meet stringent spectral requirements. For the 5th generation of mobile communication and inter-satellite communication highly integrated beam-steerable active arrays consisting of a large number of PAs and antennas are considered as an efficient solution to fulfill the new requirements [1].

Microwave power amplifiers can achieve a higher efficiency in terms of transmitted power vs. supplied power if driven as close as possible to the saturation point [2]. Unfortunately the more the PA approaches saturation, the more it behaves nonlinearly and it does not fulfill the linearity requirements dictated by the mobile communication standard. Therefore a linearity and efficiency trade-off arises which RF engineers have to deal with if they want to have the PA working with a reasonable efficiency. Back-off strategies can be used to avoid nonlinear effects: the dynamic range of the amplifier's input signal is shifted down to a lower power level so that the amplifier's output is not severely distorted by the nonlinear behavior. A solution to avoid the drastically low power efficiency in PAs using back-off is the recourse to a linearization technique. Different existing linearization methods are able to reduce the nonlinear distortions while keeping the PA as efficient as possible [3]. Digital pre-distortion (DPD) has been widely used for improvement of transmitter efficiency but generally on a single power amplifier and single antenna [4]. The new generation of active arrays considered for 5G mobile communication, is using a set of highly integrated active arrays, as illustrated in Fig.1. The active array topology where the PAs are placed after the phase shifters, and just before the antennas, gives several benefits such as power dissipation in and required power handling capability of the phase shifters, reducing the output power requirements for each element and allowing small integrated devices to be used while connected to the antennas. However, the increased complexity in the active phased array also makes it difficult to have a comprehensive understanding of the key factors contributing to the total nonlinearity of the

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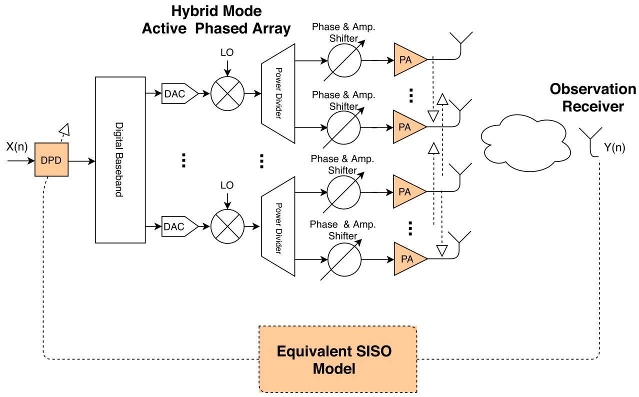


FIGURE 1. Concept illustration of the digital pre-distortion for hybrid beam-forming based on the equivalent SISO model.

whole array. Lacking this knowledge significantly limits the potential of the SISO linearization technique. The challenges responsible for the situation are as follows:

1) 5G systems are using the so-called hybrid beam-forming (Fig. 1) where the number of analog RF chains is higher than the number of digital receivers (for current consumption reductions). In such case a direct digital control of each analog RF chain is not possible and the system may not directly know the output of each amplifier. So an alternative linearization method is needed.
2) The high level of integration and large number of analog chains make the placement of feedback circuits for each branch, which is used for single antenna DPD, very challenging.
3) The signal bandwidth is rapidly increasing and it will place enormous demands on bandwidth of the analog feedback receiver used for linearization as well as sampling rate of the analog to digital converters. The feedback receiver should have a wider bandwidth than a standard receiver in order to record the distortion sidebands.
4) Using multiple transmitters and multiple antennas introduces crosstalk at PA's inputs and outputs. For mitigating the impact of these crosstalks the complexity of the algorithm is expected to increase [5] and avoiding a complexity explosion of the algorithm is another challenge.

In this paper insight into the linearization mechanisms in a 5G millimeter wave (mmWave) active phased array (APA) over the air (OTA) setup is provided through measurements. It is shown that the DPD model based on the signal captured by a single observation receiver in the far-field is able to linearize a set of PAs in an active array. The measurement set-up in this work includes an up-conversion from sub-6 GHz to mmWave which is also the general approach used by 5G manufactures in order to reuse the existing technology for mmWave. The trade-off analysis in this paper treats the following cases:

- Impact of up-conversion from sub-6 GHz into mmWave on linearization of APA.
- Impact of steering angle on trained beam in boresight.
- Comparison between a linearized SISO and a backed-off system.

This paper is organized as follows: Section I is the introduction. Section II presents active array linearization topologies in state-of-the-art solutions. Section III presents crosstalk mitigation methods. Section IV describes linearization of an active phased array. Measurement results of a 4 × 4 array as a two-port system are provided in section V. Finally, the conclusion of this work is presented in section VI.

II. ACTIVE ARRAY LINEARIZATION TOPOLOGIES, STATE-OF-THE-ART SOLUTIONS

A simple and most cost effective solution is to observe only the output of a single PA and assume that all PAs are similar which is presented in [6] and [7]. The drawback of this approach is a reduced performance.

An alternative scheme is to use an observation receiver per PA and linearize according to some averaging principle [8]. This approach is an expensive approach and requires as many observation receivers as PAs.

Another method presented in [9], [10] and [11], suggesting that the output of individual branches be combined and sampled for linearization in order to include the actual performance of each PA branch. Same approach has been suggested by [12] and [13], where they combine all branches and establish a "virtual" and "rotated" main beam. The authors show that by linearizing with a rotated observation signal, the distortion in the direction of the main beam is minimized. The method of combining the outputs requires a feedback signal from each PA output. Although this method demonstrates significant linearity improvement, it still needs bulky feedback circuits which may be impractical when using highly integrated active arrays in 5G mmWave.

Recently researches have presented the idea of SISO modelling where the entire transmitter has been considered as a two port system as illustrated in Fig. 1. We have presented a DPD technique for linearization of the antenna array in presence of crosstalk [14], using only one external observation antenna for observing the combined signal in the far field. A similar approach has been introduced by [15] and [16]. In practice, this can be implemented as part of the receiver section of the same device (i.e. diversity receiver) which has been presented by [17]. This kind of adaptive on-line OTA-DPD based on the diversity feedback uses an iterative procedure to eliminate the uncorrelated components from the feedback signal for accurate DPD. The concept has been verified by measurements with good results for small scale arrays but for large scale arrays only simulation results are available. The adaptive on-line DPD is a promising approach and needs to be investigated in more detail by industry and academia in the next years. Nevertheless, investigation of replacing the feedback antenna with a far-field observation receiver in order to analyse the impact of load modulation due to crosstalk is an important topic for the linearization approach and has been investigated in [18].

III. CROSSTALK MITIGATION, STATE-OF-THE-ART SOLUTIONS

The topologies described in II explain mainly how the response of the amplifiers is measured for DPD implementation and describe the trade-off of the systems in terms of efficiency, cost and size. The discussion about criteria by which the DPD algorithm is optimized in order to mitigate the crosstalk is another important topic. Crosstalk as coupling from one branch to another, in transmitters in an active antenna array can be categorized as two types: before PA and after PA. This is mainly due to RF leakage through the common local oscillator or coupling between different transmit paths because of the electromagnetic coupling, respectively [19].

When no isolators are present at the PAs' outputs then the PAs get a direct impact from antenna mismatch and the mutual coupling between the antennas [20]. For most practical configurations, mutual coupling is difficult to predict analytically but must be taken into account because of its significant contribution [21]. As a consequence of antenna mismatch and mutual coupling at the PA to antenna interface, signals will travel in both directions and the RF behavior of the PA must be described by a dual input dual output behavioral model. Fig. 2 shows an RF beam former including L antenna elements where a_{1k} is the incoming signal to the amplifier, b_{2k} is the output from the amplifier and a_{2k} is the reflected signal from the antenna array at the k'th branch.

Since the array elements are electromagnetically coupled, the waves fed to the antennas are also coupled back to the output ports of the PAs. This effect creates an apparent variable load at the output of each PA, depending on the operation of the transmitter [22]. Then the effective reflection coefficient (Γ_k) for the k'th element is given by:

$$\begin{bmatrix} a_{21} \\ \cdots \\ a_{2k} \\ \cdots \\ a_{2L} \end{bmatrix} = \begin{bmatrix} s_{11} & \cdots & s_{1k} & \cdots & s_{1L} \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ s_{k1} & \cdots & s_{kk} & \cdots & s_{kL} \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ s_{L1} & \cdots & s_{Lk} & \cdots & s_{LL} \end{bmatrix} \cdot \begin{bmatrix} b_{21} \\ \cdots \\ b_{2k} \\ \cdots \\ b_{2L} \end{bmatrix} \tag{1}$$

$$a_{2k} = \begin{bmatrix} s_{k1} & \cdots & s_{kk} & \cdots & s_{kL} \end{bmatrix} \cdot \begin{bmatrix} b_{21} \\ \cdots \\ b_{2k} \\ \cdots \\ b_{2L} \end{bmatrix} \tag{2}$$

$$\Gamma_k = \frac{a_{2k}}{b_{2k}} = \frac{\sum_{i=1}^{L} S_{ki} b_{2i}}{b_{2k}} \tag{3}$$

where elements in the b_2 vector are the complex coefficients describing the input to the antenna and elements in the S_k vector are the complex coefficients describing the relationship between a_{2k} and the output signal b_{2k} and determined by the characteristics of the antenna array. Since (Γ_k) is not only dependent on the reflection from the k'th element but also on the coupling from other elements in the array, then the output impedance of the k'th element is changing and as a result the linear and nonlinear behavior of it, the so-called load modulation.

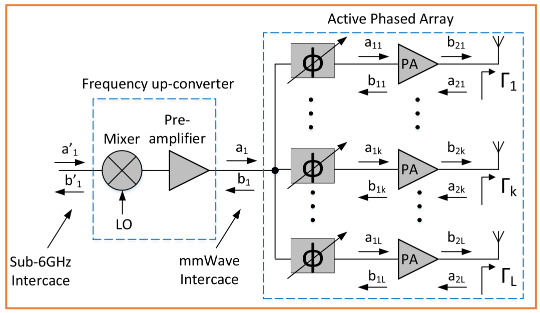


FIGURE 2. Conceptual illustration of mutual coupling and reflection coefficients of system model.

Furthermore, the behavior of each PA cannot be fully described solely as a function of its input, it will change according to the coupled signal. The input signal a_{1k} is ideally a phase shifted version of the input signal a_1. But due to gain variation of the phase shifters over phase shift setting and impact of the reflected signal b_{1k}, each PA can be driven at a different input levels for different steering angles [22].

To account for both the load modulation and steering angle dependency for each PA in the array, the dual-input PA model has been introduced in [23] and [24] where both signals, a_{1k} and a_{2k} are included in the nonlinear function. These works demonstrated good results by applying DPD to the beam forming array. However, the method requires that the s-parameters for the antenna are known and it needs feedback from each antenna. The high number of PAs and the compact size of modules makes this approach very challenging.

We have in [14] presented a system level SISO DPD technique for linearization of the antenna array in presence of crosstalk and similar work has been presented in [15]. The reported measurement results are limited to implementations at sub-6 GHz and small array sizes, and challenges specific to DPD when applied to mmWave arrays still remain.

As mentioned in section I, a general approach used by 5G manufactures is to use frequency up-conversion from sub-6 GHz. As illustrated in Fig. 2 the input signal to the APA, a_1, is ideally a frequency up-converted part of the sub-6 GHz signal a'_1. But in reality there would be mismatch on output and input of the frequency up-conversion block due to reflected signals, b_1 and b'_1, as well as distortion in the up-conversion mixer and pre-amplifier. Including the impact of these blocks to the system model makes the DPD algorithm more complex. In this work the impact of these blocks to the applied DPD is evaluated through measurements.

IV. LINEARIZATION OF ACTIVE PHASED ARRAY

In this section the linearization method using the radiated far-field signal of an active array at a single observation

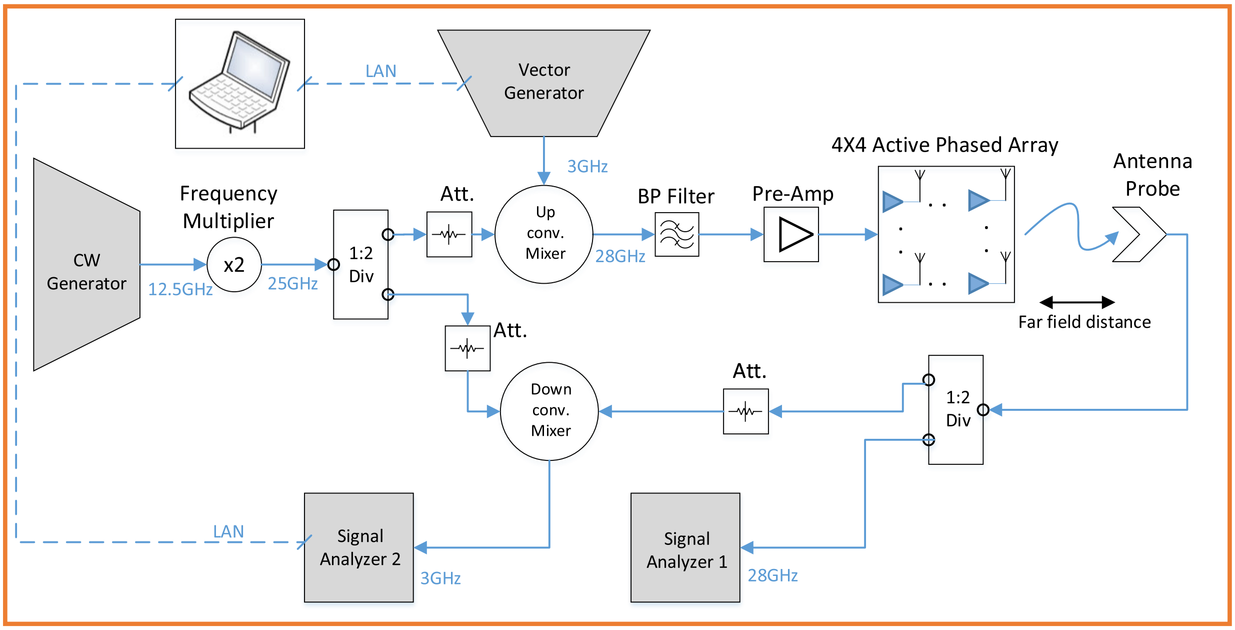


FIGURE 3. Block diagram of OTA SISO memory polynomial model based digital pre-distortion for the 28 GHz active phased array.

receiver is described. The assumption is that the nonlinearity of the PA is the main source of distortion and not the crosstalk. Equation (4) represents the applied memory polynomial model (MPM) which is a deviation of the Hammerstein model and has been proven effective for removing nonlinearity and memory effect [25]:

$$y(n) = \sum_{k=1}^{K} \sum_{m=0}^{M} a_{km} x(n-m) \left| x(n-m) \right|^{k-1} \tag{4}$$

where a_{km} is the 2-D array of filters and power series coefficients of the amplifier, K is nonlinearity order of the memory polynomial and M is the highest memory depth.

Since the a_{km} coefficients are linear weighting of nonlinear signals then these coefficients can be found using the least-square type algorithm. The easiest way to formulate such an algorithm is to first collect the coefficients in a J × 1 vector denoted ω, where J is the total number of coefficients. Then the model output can be expressed using the following equation written in vector form.

$$\tilde{y} = P\omega \tag{5}$$

where:

- $\tilde{y}$ is a N × 1 vector representing an estimate of the amplifier actual output.
- P is a N × J matrix where N is the number of samples and J is equal to M times K.
- ω is a vector with J × 1 coefficients.

The inverse of this model used for pre-distortion is then:

$$\tilde{x} = Rw \tag{6}$$

where R is defined similarly to P now with $y(n-m)$ replacing $x(n-m)$ in equation (4). The input is now estimated from the output samples and the estimation error can be calculated as:

$$e = x - \tilde{x} \tag{7}$$

The best estimate for getting a_{km} coefficients is to use a least square solution which minimizes the squared error:

$$w = (R^H R)^{-1} R^H x \tag{8}$$

In this work the parameter extraction was performed using the Moore-Penrose pseudo-inverse because this technique provides a more robust solution to the system and avoids instability in parameter extraction due to the eventually high condition number of the model matrix [3].

V. MEASUREMENT RESULTS OF 4 × 4 ARRAY AS TWO-PORTS SYSTEM

A. MEASUREMENT SETUP

The block diagram of the measurement setup for the 4 × 4 array is shown in Fig. 3 and the actual measurement set-up is illustrated in Fig. 4. The input source for the measurements is a 3 GHz LTE10 signal, compliant with the 3GPP downlink orthogonal frequency-division multiplexing (OFDM) modulation with a peak to average power ratio of 10.6 dB from the signal generator. For up-conversion, an unmodulated signal of 12.5 GHz has been frequency doubled to 25 GHz and fed into a power divider in order to be used as local oscillator (LO) signal for both up-conversion and down-conversion. A 28 GHz band-pass filter is used to select the up-converted modulated signal and suppress the LO leakage and image frequency signals. In order to avoid any nonlinearity in the multiplier and up-converter, the signal levels in these stages are kept in the linear operating ranges of these devices according to their specifications and then amplified by a pre-amplifier in

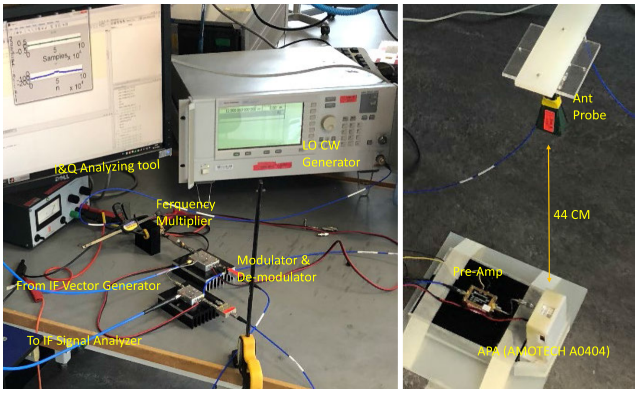


FIGURE 4. **OTA SISO measurement setup of the 4 × 4 active phased array; using up-conversion from 3 GHz to 28 GHz.**

order to reach the level necessary for driving the active array into compression. The pre-amplifier has a gain of 30 dB and at output levels above 8 dBm (including cable loss), it deteriorates the linearity of the signal which is shown in the measurement result in next section. The 28 GHz signal is fed to an AMOTECH A0404 which includes four Anokiwave AWMF-0158 [26]. This device integrates 16 branches of attenuators and phase shifters plus PAs and 16 patch antennas in a 4 × 4 active phased array.

The diameter of the active array antenna is approximately 4 cm which at 28 GHz results in to a far-field distance of:

$$\frac{2D^2}{\lambda} = 30.5\ cm \tag{9}$$

where D is the diameter of the antenna and λ is the wavelength. The main beam signal is captured by the observation horn antenna placed 44 cm away which is well above the far-field distance of the device.

Problems with reflections from the surroundings are not observed since the distance between the active array and the horn antenna is short, so the reflected signal from the rack and the set-up are much weaker than the desired signal. Furthermore the active array and horn antenna have very good directivity so reflections are expected to be well attenuated. Therefore measurement without an anechoic chamber is expected to be no issue.

The captured signal is split into two branches in order to both be analysed with the signal analyzer for monitoring the actual adjacent channel power ratio (ACPR) at 28 GHz and be down-converted to a 3 GHz signal and captured by another signal analyser for getting access to I and Q data. The input power from the signal generator is adjusted in order to get an root mean square (RMS) level up to 8 dBm into AWMF-0158 which according to [26] drives the PAs into compression.

The steps of the experiment are:

1) I and Q data of the modulated signal from the vector signal generator were acquired using a sample rate of 100 MHz at the signal analyzer and recorded by a computer.
2) The recorded I and Q data are loaded into the vector signal generator, generating a modulated signal at 3 GHz which is then up-converted to 28 GHz. The active array is excited using this modulated signal and the output, $y(n)$, is captured at the observation point by the receiver antenna probe. This signal is then down-converted and acquired by the signal analyzer and the I and Q data recorded into the computer.
3) The input and the recorded output signals are up-sampled to a finer resolution, time-aligned using cross-correlation, down-sampled and then used for the predistorter identification.
4) The pre-distorted signal is now generated by the memory polynomial with a memory depth of M = 8 and a linearity order of N = 5 using recorded $x(n)$ and $y(n)$.
5) Once the predistorter is identified, the signal is uploaded to the vector signal generator and again applied to the APA. The corresponding output signal is recorded and the power spectral density (PSD) and error vector magnitude EVM) are calculated based on the recorded I and Q samples.
6) Output power and ACPR of the 28 GHz signal are directly measured by the signal analyzer.

TABLE 1. **EVM and ACPR measurement results without and with DPD.**

Measurements	EVM[%] (wo/w)	ACPR[dB] Lower(wo/w)	Upper(wo/w)
case 1	6.21/5.10	-40.03/-43.84	-40.11/-43.53
case 2	6.21/3.96	-38.27/-45.19	-38.08/-44.94
case 3	7.20/5.84	-36.60/-42.48	-36.69/-41.87

B. MEASUREMENT RESULTS

The measurement has been done in 3 cases: APA input at 6 dBm, 8 dBm and 10 dBm respectively. According to the APA data sheet it has a 1 dB compression point at around 8 dBm input. Table 1 shows the EVM and ACPR measurement results with and without DPD for these 3 cases. The best result is achieved in case 2 where ACPR and EVM are improved by respectively 7 dB and 2.3 % using the applied DPD. As expected the linearity improvement for case 1 with 6 dBm input is not as good as case 2 because the APA is not enough in compression. Fig. 5 and Fig. 6 show the measurement results of PSD and amplitude to amplitude (AMAM) distortion gain for these 3 cases. The small linearity improvement in case 3 could be explained due to nonlinearity of pre-APA blocks.

C. IMPACT OF UP-CONVERSION FROM SUB-6 GHz INTO mmWave ON LINEARIZATION OF APA

The linearity of the up-conversion blocks are expected to have an impact on SISO DPD since the nonlinearity could partly be from a pre-APA block if it is in compression.

To investigate the pre-APA blocks nonlinearity, the ACPR of the system after the up-conversion mixer and after the

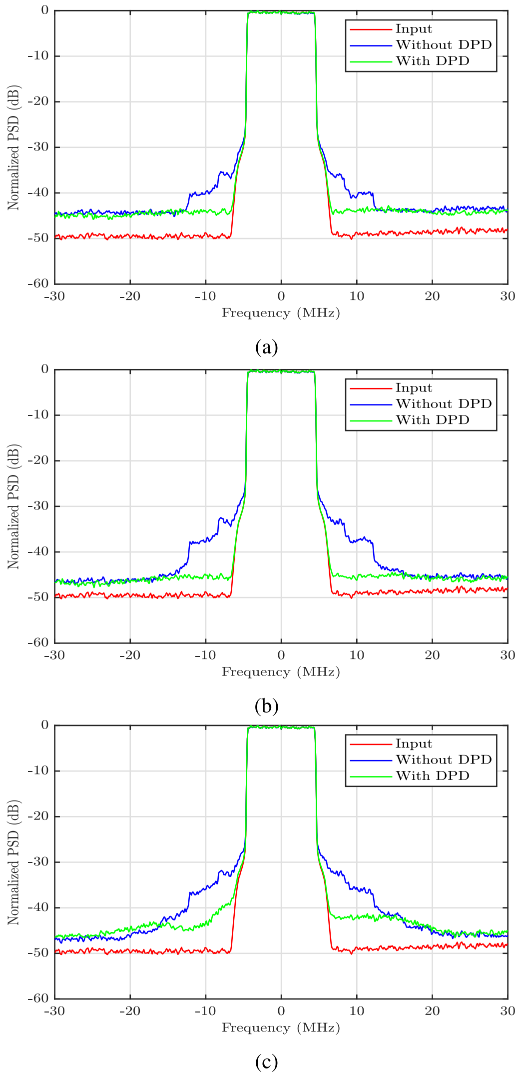


FIGURE 5. Measured power Spectral density with and without pre-distortion: (a) case 1 with input power of 6 dBm; (b) case 2 with input power of 8 dBm; (c) case 3 with input power of 10 dBm.

TABLE 2. EVM and ACPR measurement results at the up-conversion Mixer's output.

Measurements	EVM[%]	ACPR[dB] Lower	ACPR[dB] Upper
case 1	0.89	-46.06	-46.35
case 2	0.89	-46.68	-47.01
case 3	0.89	-47.01	-47.50

TABLE 3. EVM and ACPR measurement results at the pre-amplifier output.

Measurements	EVM[%]	ACPR[dB] Lower	ACPR[dB] Upper
case 1	0.89	-42.23	-42.29
case 2	0.90	-40.51	-40.52
case 3	0.89	-38.50	-38.60

pre-amplifier are measured and the PSD plots are shown in Fig. 7 and Fig. 8 respectively and in tables 2 and 3 the EVM and ACPR results are listed.

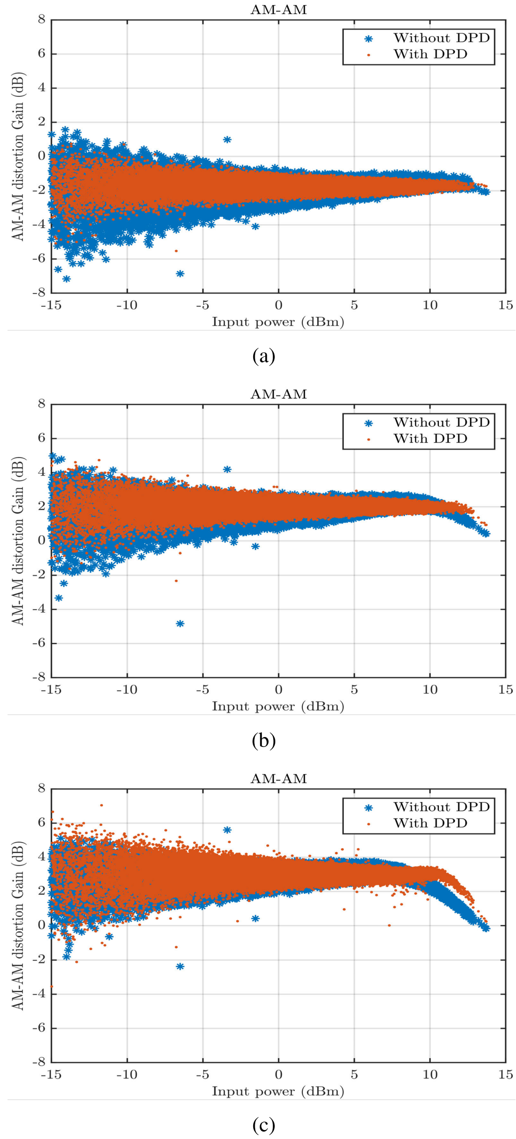


FIGURE 6. Measured AM-AM distortion gain with and without pre-distortion: (a) case 1 with input power of 6 dBm; (b) case 2 with input power of 8 dBm; (c) case 3 with input power of 10 dBm.

The results indicate that the signal is linear after the up-conversion mixer but the pre-amplifier is running into high compression in case 3, and as a consequence the ACPR is increased. The applied DPD is not capable to mitigate the nonlinearity caused by pre-amplifier in case 3 properly. One reason could be a time delay between pre-amplifier and APA which needs further investigations.

D. IMPACT OF BEAM ANGLE

In section III the impact of mutual coupling between the antennas at PAs' outputs has been discussed and it was mentioned that the impedance presented at the output of each PA depends on the mutual coupling between antennas. In this section this impact is proven through measurements. The procedure is as follow: the placement of the observation receiver antenna has been kept fixed at maximum received

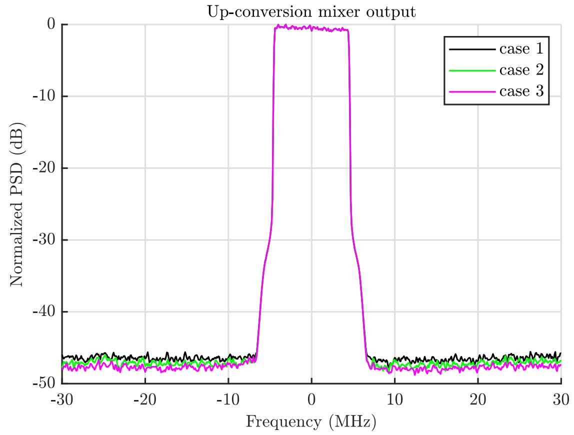


FIGURE 7. Measured ACPR at the up-conversion Mixer out: case 1 (black) corresponds to 6 dBm at the APA input; case 2 (green) corresponds to 8 dBm at the APA input; case 3 (magenta) corresponds to 10 dBm at the APA input.

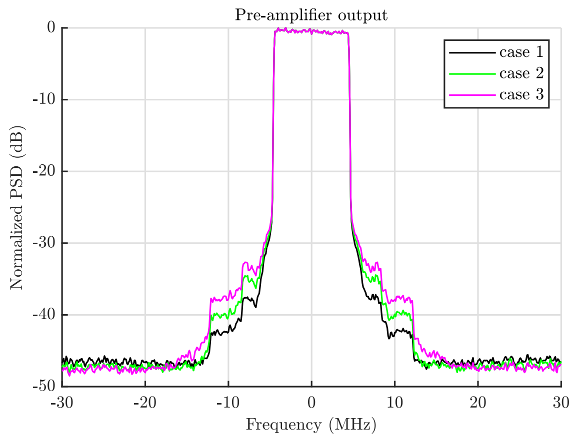
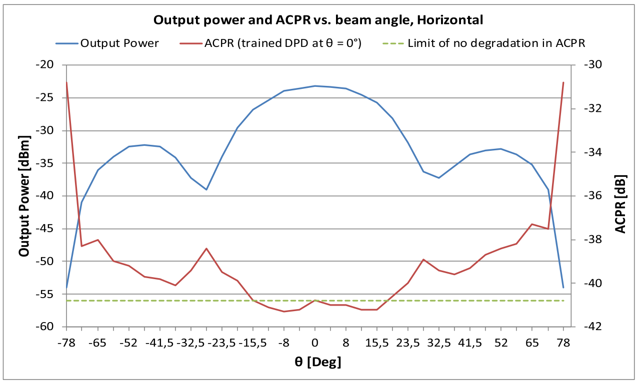


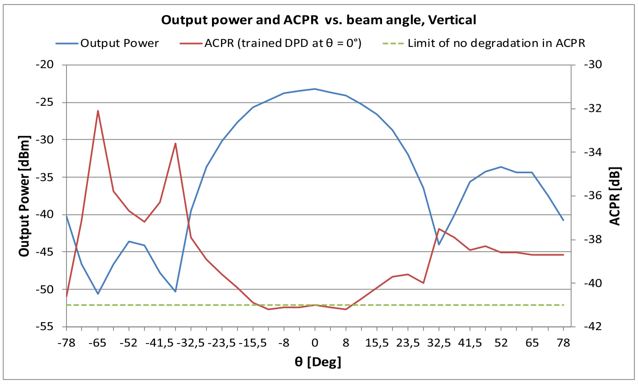
FIGURE 8. Measured ACPR at the pre-amplifier out: case 1 (black) corresponds to 6 dBm at the APA input; case 2 (green) corresponds to 8 dBm at the APA input; case 3 (magenta) corresponds to 10 dBm at the APA input.

signal ($\theta = 0$ degree). The DPD has been trained and the pre-distorted input has been detected at this position. Then while using this pre-distorted input, the main beam of the APA has been shifted from $\theta = -78$ to $+78$ degrees in approximately 5 degrees step using the code-book and software tools of AMOTECH A0404. The measurement is done for both horizontal and vertical steering angles of the main beam. The measurement result is shown in Fig. 9. The magnitude of the beam captured by the fixed antenna probe is varying by changing the beam direction as expected. Fig. 9b shows the magnitude and ACPR on the left side are worse than on the right side. This is due to the asymmetrical structure of the actual AMOTECH A0404 device where the placement of the connector has an influence on the beam and is not a general issue.

However a single trained DPD is not sufficient for maintaining a low ACPR in a wide range of steering angles. To maintain an ACPR level below 41 dBc across the steering angle, a new training after approximately ± 15 degree shift of the main beam is required. This can be explained as the effect of mutual coupling of the highly integrated antennas in the array and due to variation in input levels because of gain variation of phase shifters. Fig. 9 also shows that there is a symmetry of the array and the trained DPD at $+\theta$ can be used for a steering angle of $-\theta$ which reduces the number of training steps required for linearization across the entire steering range.



(a)



(b)

FIGURE 9. Impact of beam angle on linearization, (a) horizontal beam steering, and (b) vertical beam steering.

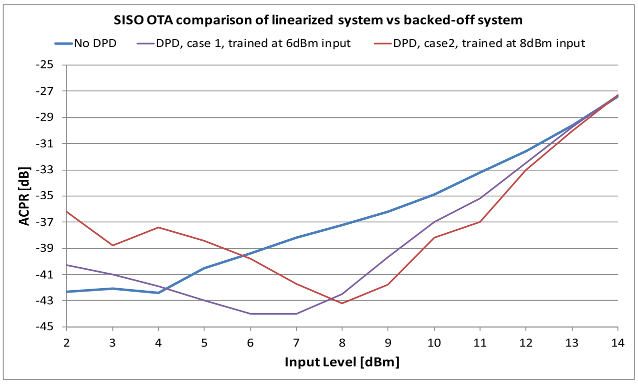


FIGURE 10. Comparison of a linearized system to a backed-of system, in terms of ACPR, at varying input back-off levels.

E. COMPARISON BETWEEN LINEARIZED SISO AND BACKED-OFF SYSTEM

In order to minimize the complexity it is desired to keep the trained pre-distorted signal as input for a range of output powers if possible. At the same time it is desired that the linearized system has not degraded performance compared to a backed-off system around the back-off point. To investigate this, the trained pre-distorted signals in case 1 and case 2 of section V have been reused for a set of output powers of ±6 dB and the resulting ACPRs are compared with the case without pre-distortion. As it is shown in Fig. 10, the linearity compared to a backed-off system, is always better for output powers above the trained point and down to 2 dB below this point.

VI. CONCLUSION

In this paper the SISO OTA linearization trade-offs of a 4×4 active array running at 28 GHz, modulated with an LTE10 DL OFDM signal have been investigated. The results indicate:

i. there is a trade-off between the complexity of the applied DPD algorithm and the linearity of the pre-APA blocks under test. Increasing the gain of these pre-APA blocks, which can affect their linearity, will limit the capability of a less complex DPD algorithm. With the applied OTA DPD, up to 7dB improvement of ACPR and 2.3 % improvement of EVM are achieved with minimum complexity of the algorithm.

ii. the linearized beam is sensitive to beam angle. The trained beam cannot be reused for beam angles above a certain limit which could be explained due to mutual coupling and crosstalk between antennas and due to variation in input levels. A new trained beam is required in order to avoid ACPR degradation when changing beam steering angle. In this investigation, for maintaining the ACPR level across the steering angle, a new training after approximately ±15 degree shift of the main beam is required.

iii. near the back-off region, reusing a set of trained coefficients, the linearized system is out-performing the backed-off system from above the trained output level and down to a certain level below that. The achieved result in this work is 2dB below the trained level.

Future work based on the presented SISO OTA technique may include complexity analysis of the DPD algorithm in the case of distorted pre-APA blocks and the impact of channel properties on the SISO OTA linearization.

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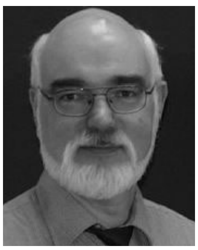

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